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THE SELECTIVITY AND INTERMODULATION PROBLEM IN UHF COMMUNICATION EQUIPMENT*

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The problem of intermodulation is not new. With the increasing use of uhf and vhf and the attendant close channel spacings, it is merely emphasized. Despite the fact that it is a very old problem, a casual review of the old reliables in the handbook and textbook fields shows that a few of the books mention the word "intermodulation"; none give it any attention. In fact, one would gather the impression that the subject is unimportant. The purpose of this paper is to review the subject of intermodulation, with particular emphasis on the elements of the problem at ultra high frequencies. In the presentation, I am sure, you will find nothing new. The purpose of the talk is merely to describe some of the methods which can be employed to reduce intermodulation, and their advantages and disadvantages. As you doubtless know, intermodulation is not a serious problem at low and medium frequencies because adequate protection in the way of receiver preselection can be achieved. Likewise, I am sure, you have experienced cases of external intermodulation where spurious signals are produced by rectification which occurs at poor or corroded junctions in current carrying systems of wires, steel structures, and the like, which are located near the receiving antenna. At vhf and uhf, external cross modulation, while it is occasionally observed, is not the principal offender. The principal offender is the receiver itself. Why then, you may ask, isn't something done about it? It can be safely stated that much has been done, but we can also add that there is much that remains to be done if the maximum use of the closely spaced vhf and uhf channels is to be obtained.

You will recall that the intermodulation terms which are of interest are of the general form $2f_a - f_b$, or even more generally $f_a + f_b - f_c$. Fig. 1 shows only a few of the possibilities. The $2f_a - f_b$ possibilities are quite clear; those of the type $f_a + f_b - f_c$ are not quite as easy to pick out. The simplest guide that can be given is that, if the frequency difference between two channels produces a beat which, when considered as a modulating voltage can modulate a third frequency and produce a sideband on the desired frequency, it is an interfering combination. For example, take channels located beyond the chart shown, say channels 8 and 13. The difference frequency is five times channel spacing, so that channel +5 or -5 will be an interfering combination. An example on the chart could be channel +2 and channel -3; difference frequency, 5; trouble ahead if either channel +5 or -5 is active in the area.

To date, the principal offenders have been of the type $2f_a - f_b$. As channel utilization is increased, the trio problem will rear its ugly head. As an example, suppose channels 6,7,8,9,10, and 11 are operating simultaneously in one location, and we operate on channel zero at a point nearby, say the same town. A new facility is to be established, and a search is

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made for a new noninterfering channel to be activated in the area. In the search for a noninterfering channel, we can begin by ruling out channels that will produce interference on channel zero. Hence, channel +5 is out because of operation on channels 6 and 11. Likewise, 4,3,2,1,-1,-2,-3,-4, and -5. The first channel that will not produce interference with channel zero on the low side is channel -6. On the high side, the first operable channel is channel +16. Thus, the assignment of seven channels in the area, on a pattern in accord with the foregoing example, vitiates the assignment of an operating frequency on any channel -5 through +15. Twenty channels blocked because of intermodulation! This is not to say that legal authorization could not be obtained, but rather that operation of any of the channels in the band -5 to +15 would be subject to intermodulation or co-channel interference.

So far, I have only indicated the possibility of the existence of intermodulation and the relationships between offending frequencies. The first step in our further consideration has to do with the magnitude of the interference produced. It is, of course, obvious that the near perfect answer to our problem is a high degree of preselection, so that the only signal reaching the grid of the first vacuum tube is the desired signal, and an adequate amount of attenuation is provided for the adjacent channels and all channels farther removed. In the vhf field, where sensitivity is a prime consideration, ordinary set designs provide something of the order of 50-db protection from intermodulation, and some carefully produced equipments may reach 60 to 70 db of protection. This figure is determined as shown in Fig. 2. The receiver under test is equipped with any kind of a signal level indicator -- in AM receivers, an AGC voltage reading, or a signal-to-noise reading, and in FM receivers, a limiter reading or a quieting reference.

A simple substitution arrangement can then be used, and the magnitude of protection against intermodulation determined. It is both convenient and desirable, as well as standard practice, to keep the output of the ΔF and $2\Delta F$ signal generators equal, and the magnitude of the cross modulation protection is the ratio of undesired to desired signal power that results in equal receiver response. This ratio is given in decibels.

A simple extension of the experiment will show that the interfering signal is proportional to the square of the output of the ΔF signal generator and directly proportional to the output of the $2\Delta F$ signal generator. Fig. 3 shows this relationship. The curve shown is for a 50-db protection. It is particularly important to note that as the output of the ΔF signal generator is increased, the signal output of the $2\Delta F$ generator can be reduced to very low values. For example, when the ΔF generator output is 75 db above 1 microvolt, a signal of 1 microvolt at $2\Delta F$ will produce a response equal to the 1-microvolt desired signal. Consequently, in practice, a strong nearby station, operating a few channels removed in the vhf band, can cause interference when combined with a quite weak signal twice that number of channels removed. I have shown a case where a 50-db intermodulation protection is given. This is not an impractical case, and too many vhf and uhf receivers are afflicted with this shortcoming. And, I might add that additional db's of protection do not come easily.

It is quite true that the use of a high degree of selectivity before the received energy is applied to the input terminals of the first vacuum

tube in the receiver system, will result in the reduction of intermodulation. It is appropriate to examine a few of the problems associated with the achievement of RF selectivity at vhf and uhf.

Fig. 4 shows the 3-db bandwidth of a single tuned circuit as a function of its unloaded Q, and the amount of permissible insertion loss. Remember that at vhf and uhf the vacuum tube grid circuit is a dissipative element and power must be delivered to the grid input circuit. As a result, we must consider the factors, loaded bandwidth and permissible loss, which are generally of no importance at medium and high frequencies. When the receiver designer is striving for maximum receiver sensitivity, a low insertion loss is a requirement. The curve shows the effect of such a requirement on the loaded bandwidth. Note that, if the unloaded Q of the circuit is 1,000 (not a bad figure), that if the insertion loss is to be held as low as 0.6 db, the unloaded bandwidth is approximately 1 mc. This, at 150 mc. At 300 mc, equivalent figures yield 2 mc, or 20 of the 100-kc channels! Why not, you ask, use a cascaded preselector? This path leads to more insertion loss and provides skirt selectivity, but does not greatly help in the neighborhood of the desired carrier. To show this, consider the following problems:

Operation frequency.	225 mc
Channel width.	0.1 mc
Loaded Q.	2,250
Unloaded Q for a single resonator with 0.6-db insertion loss. .	18,000

This single circuit would provide an adjacent channel rejection of 7 db. The response across the desired band would vary 3 db. Additional tuned circuits would improve the adjacent channel rejection, but at the expense of 0.6 db of sensitivity per tuned circuit added.

Fig. 5 shows this relation, up to n = 4 circuits. In this example, we have mentioned, without comment, resonators with an unloaded Q of 18,000.

Fig. 6 shows an estimate of unloaded Q as a function of the volume of the resonator, and without attempting to say much about its form. Probably, an electrodynamicist could derive a formula concerning the maximum Q obtainable in a given volume at a given frequency, but the curve shown is pretty much in line with experience. Obviously, we do not put circuits of Q = 18, 000 in receivers. Volumes of 1 cubic foot per tuned circuit are not too readily accepted by the customer, particularly the military.

In cases where severe intermodulation problems have been encountered, however, coaxial cavities are used in fixed stations. Fig. 7 shows a 25- to 40-mc cavity and a 150- to 180-mc cavity. A sectional drawing of the 150-mc cavity is shown in Fig. 8. These units, as a matter of fact, have found wide use at fixed stations, and the loaded Q is adjustable by means of interchangeable coupling loops. The design includes temperature compensating features necessary when extremely high Q operation is mandatory.

So much for preselection. Let us next direct our attention to the vacuum tube, the nonlinear culprit in the system. The plate current response of a vacuum tube, ideally, is proportional to the three halves power of the average voltage at the grid plane. Practically, deviations from the 3/2 power law are experienced due to a variety of reasons, among them the initial emission velocities, intentional or unintentional nonuniformity of

grid wire spacing, and others. Irrespective of these deviations, the plate current of a tube, say a pentode, for simplicity's sake, can be expressed as a power series where the independent variable, the grid voltage, is expressed in terms of its deviation from the static bias value. When a power series expansion is employed, it is not too difficult a problem to determine the magnitude of intermodulation effects through a straightforward mathematical procedure. Fig. 9 shows the result of such a consideration. This treats the condition existing in a low level RF amplifier stage. It likewise shows the computed coefficients of the power series expansion for a 6BH6 tube operating at a screen voltage of 130 volts and a bias voltage of -1.5 volts. The computation shows that equal signals on adjacent and alternate channels some 85 db above 1 microvolt produce an on-channel interference signal of 1 microvolt. This figure of 85 db is the basic intermodulation protection for a desired signal of 1 microvolt. One element not particularly evident in this figure should be brought out. That is, if the adjacent and alternate channel signals are increased 10 db, the interfering signal, due to intermodulation, is increased 30 db. The intermodulation protection is consequently reduced from the figure of 85 db to 65 db as a result.

Fig. 10 shows the problem of the heterodyne converter. In this case, we have taken as an example a 6BH6 mixer operating at a screen voltage of 130 volts and a grid bias voltage of -3 volts. The coefficients in the power series expansion are shown, and the computed intermodulation protection for this mixer, operating with a local oscillator voltage of 1 volt applied, is 77 db. This intermodulation protection is that which exists at a desired signal level of 1 microvolt. The relationship between the mixer case and the amplifier case is similar in form, the difference being the particular power series coefficients appearing in the expression.

At this point, we might stop to consider the situation which exists in a receiver having one RF amplifier stage prior to the first mixer. If the RF amplifier stage has a gain of 10 db, the adjacent and alternate channel signals applied to the mixer are practically 10 db higher than the signals applied to the RF amplifier stage. A cascade of this type, using the values shown in Figs. 9 and 10, would show that intermodulation protection at an input signal of 1 microvolt would be 85 db in the amplifier stage and 77 minus 20, or 57-db intermodulation protection in the mixer stage. This statement assumes that the degree of selectivity achieved in the RF system is not of any substantial magnitude at the adjacent and alternate channel frequencies. In the 150-mc band, and even more so at the 225-400-mc military band, this is quite true. One particular point can be seen in the consideration of these figures; that is, for optimum receiver design and maximum intermodulation protection, signal levels at the uhf mixer grid should be held as low as is consistent with signal-to-shot-noise considerations. You may well inquire, why use an RF amplifier stage at all? The answer: the achievable signal-to-noise ratio in mixers is not as favorable as can be obtained in straight RF amplifier stages. The basic conclusions that must be drawn from any consideration of intermodulation can be made at this point; namely, receiver gain up to the mixer grid must be held to a minimum consistent with signal-to-noise considerations, and then, after conversion, adequate selectivity elements can be employed. After the signal passes through the single channel selectivity element, gain can then be added without penalizing the intermodulation performance of the

receiver. This principle of receiver design has been strictly adhered to in the Motorola Sensicon line of two-way uhf radio equipment.

Fig. 12 shows a photograph of such a receiver. The rectangular box in the receiver chassis contains a sealed, potted selective filter. The principal gain of the receiver is achieved after the selection of the desired channel through the use of this filter. Receivers of this type have been produced with provide intermodulation protection in excess of 65 db. This figure is achieved through the use of high Q coaxial resonators in the RF chain and the careful selection of the operating points of the amplifier and mixer tubes. In addition, carefully controlled RF gain limits have been established.

Voltage magnitudes alone are hardly adequate to emphasize the importance of intermodulation protection and control. An example of its effect will better serve to illustrate the magnitude of the effect. First, assume that a receiving station is established, equipped with a dipole antenna located at a height of 100 feet above ground. Further assume that the receiver provides an intermodulation protection of 50 db at its operating frequency of 150 mc, with the intermodulation figure based upon an equivalent 1-microvolt signal on its 50-ohm input cable. At a distance of 5 miles, an adjacent channel transmitter having a power output of 250 watts into a dipole 200 feet above the ground is licensed to operate. Under these conditions, what alternate channel transmitter operation will produce an equivalent interfering signal of 1 microvolt on the receiver cable? The answers are:[1]

1. 20-watt airborne transmitter, altitude 10,000 feet, all distances within 160 miles.
2. 250-watt ground station, antenna height 200 feet, any distance less than 54 miles.
3. 10-watt mobile station, antenna at ground level, all distances less than 11 miles.

These, of course, are ideal answers, over good soil, flat country. The importance of such figures can scarcely be underestimated. A 10-db improvement in intermodulation provides the following relief:[1]

Airborne example — reduction from 160 to 90 miles
Ground station example — reduction from 54 to 21 miles
Mobile station example — reduction from 11 to 2.2 miles.

A further improvement of 10 db lowers the figures to 21, 4.6, and 0.4 miles, respectively.

The foregoing consideration shows that interference can exist due to the operation of stations located at points quite remote from the receiving station. It can be expected that both two and three station combinations will prove troublesome in concentrated areas such as large cities, air bases, and the like.

To summarize the situation, we can say that intermodulation can be controlled through 1. careful receiver design, and 2. judicious use of external cavities. Obviously, in many applications, external cavities are not

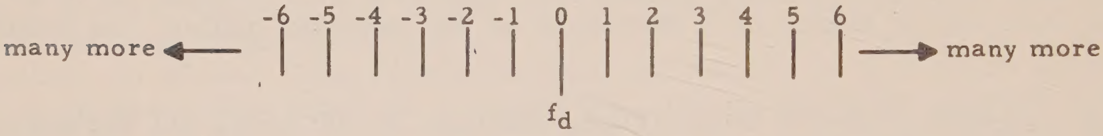
feasible at 30 mc, somewhat more feasible at 150 mc, and are possibly a very practical accessory in the 250- to 400-mc band.

At the beginning of this talk, I mentioned that much had been done and much remains to be done in the control of intermodulation in uhf re-receivers. I have outlined the problem, shown some of the things that have to be done, and now, you are correct in asking the question: What can be done in the future to improve intermodulation protection? Two possibilities certainly exist: 1. the development of a vacuum tube characteristic where third and fourth order curvature is minimized and 2. for fixed stations, at least, the development of harmoic mode crystal filters is a possibility. I am not underestimating the practical difficulties in achieving either of these results. However, until such time as adequate precision uhf selectivity is obtained, it is almost certain that the intermodulation problem will remain with us.

References

1. Based upon NDRC Division 15 Report No. 966-6C; October 19, 1944.

60 kc. Channel Spacing



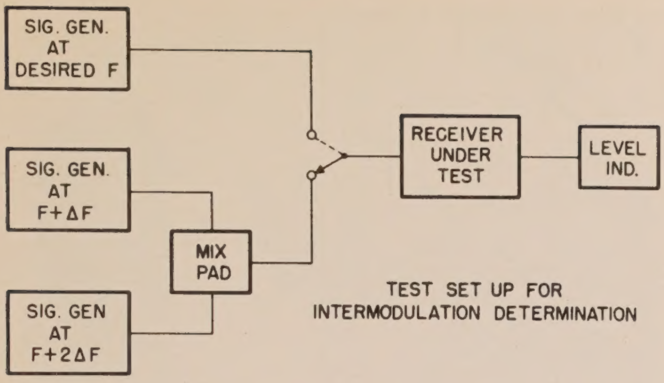
Pairs of channels that can produce a response in a receiver tuned to f_d (Channel Zero)

Channels	Response
1 and 2	2 x 1 - 2 = 0
-1 and -2	2 x -1 - (-2) = 0
+2 and +4	2 x 2 - 4 = 0
-2 and -4	2 x -2 - (-4) = 0
and so on.	

Trios of channels that can produce a response in the receiver tuned to f_d (Channel Zero)

Channels	Response
2, 3 and 5	2 + 3 - 5 = 0
3, 4 and 7	3 + 4 - 7 = 0
-2, -3 and -5	-2 - 3 - (-5) = 0
-3, -4 and -7	-3 - 4 - (-7) = 0
1, 3 and -2	3 + (-2) - 1 = 0
-1, -3 and +2	-3 + 2 - (-1) = 0
1, 3 and 4	1 + 3 - 4 = 0
-1, -3 and -4	-1 - 3 - (-4) = 0
-1, 3 and 4	-1 + 4 - 3 = 0
1, -3 and -4	1 - 4 - (-3) = 0
and many, many more	

Fig. 1 - Intermodulation at vhf (152-mc band).



TEST SET UP FOR INTERMODULATION DETERMINATION

THE RELATION:

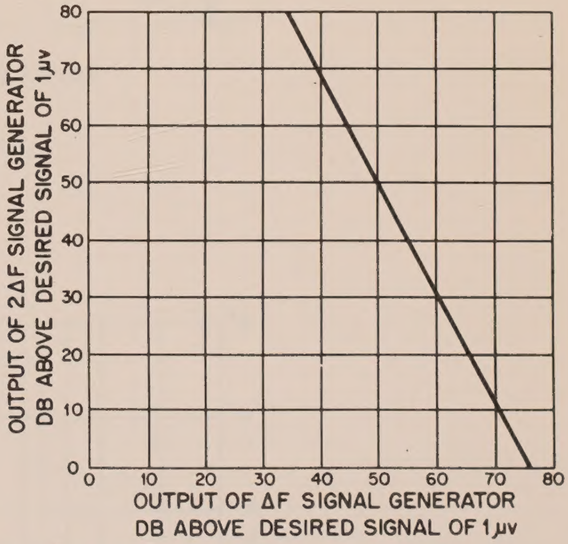
$$E^2_{F+\Delta F} \times E_{F+2\Delta F} = K^3 E_{DES}$$

20 LOG K IS INTERMODULATION PROTECTION IN DB

MORE GENERALLY,

$$E_1 \times E_2 \times E_3 = K^3 E_{DES} \quad \text{(INTERFERING } f = f_1 + f_2 - f_3\text{)}$$

Fig. 2



ΔF AND 2ΔF INTERFERING SIGNALS REQUIRED TO PRODUCE AN EQUIVALENT ONE MICROVOLT SIGNAL ON THE DESIRED CHANNEL

THE RECEIVER PROVIDES 50 DB INTERMODULATION PROTECTION

Fig. 3

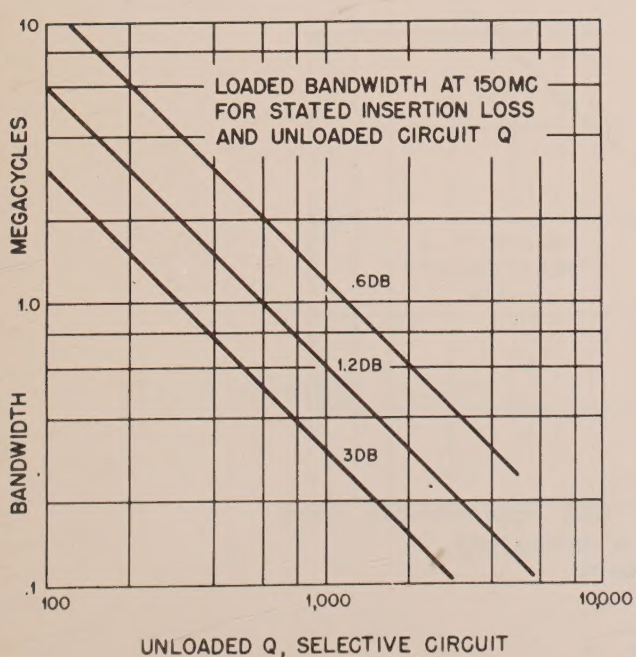


Fig. 4

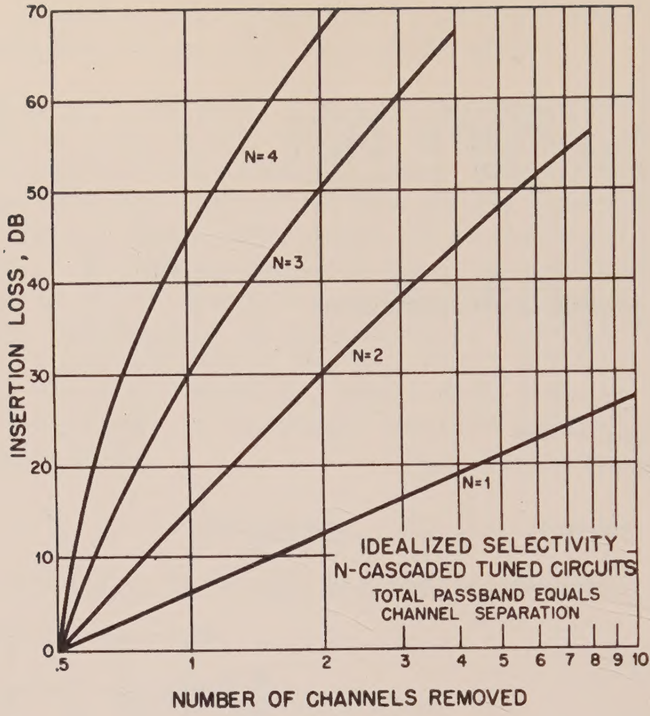

70
60
50
40
30
20
10
0
N=4
N=3
N=2
N=1
IDEALIZED SELECTIVITY
N-CASCADED TUNED CIRCUITS
TOTAL PASSBAND EQUALS
CHANNEL SEPARATION
INSERTION LOSS, DB
.5
1
2
3
4
5
6
7
8
9
10
NUMBER OF CHANNELS REMOVED

Fig. 5

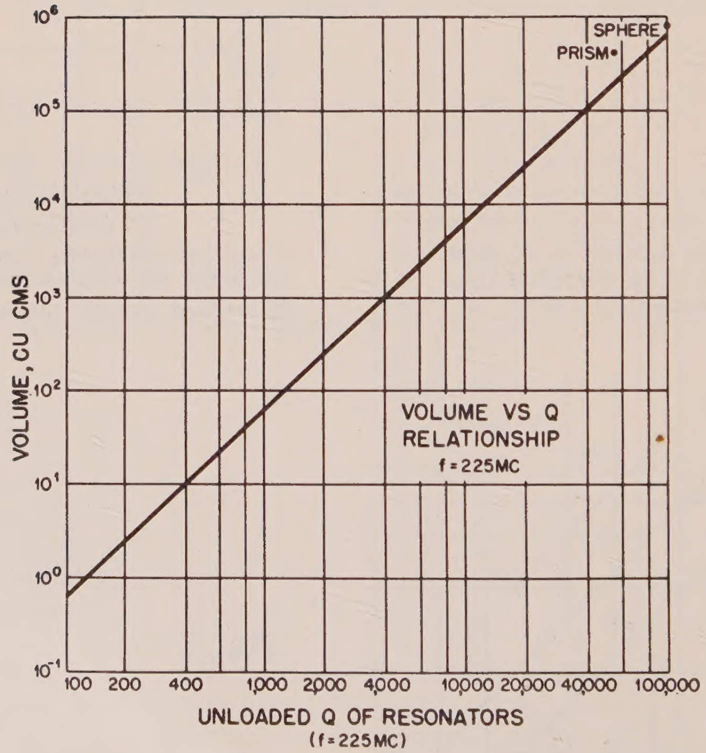

10^6
10^5
10^4
10^3
10^2
10^1
10^0
10^{-1}
SPHERE
PRISM
VOLUME VS Q
RELATIONSHIP
f = 225MC
VOLUME, CU CMS
100
200
400
1,000
2,000
4,000
10,000
20,000
40,000
100,000
UNLOADED Q OF RESONATORS
(f = 225MC)

Fig. 6

Fig. 7

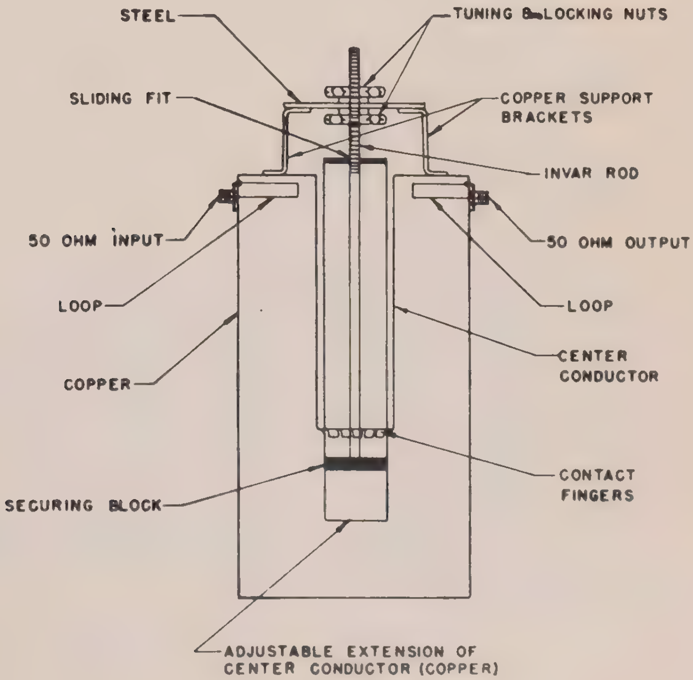


Fig. 8 - Cavity filter.

$$i_p = a_0 + a_1E + a_2E^2 + a_3E^3 + \dots$$

$$E_i = \frac{3a_3}{4a_1} E_\Delta^2 \times E_{2\Delta}$$

where
E_i = intermodulation signal
E_Δ = adjacent channel signal
$E_{2\Delta}$ = alternate channel signal

when
$E_\Delta = E_{2\Delta}$ (standard IM test)

$$E_i = \frac{3a_3}{4a_1} E_u^3 .$$

Experimental values of coefficients for 6BH6 tube (E_s = 130 V, E_c = -1.5 V):
a_0 = 2.50
a_1 = 2.29
a_2 = 0.36
a_3 = 0.427.

Computed intermodulation value: 85.6 db for 1 microvolt spurious.

Fig. 9 - RF amplifier intermodulation.

$$i_p = a_0 + a_1E + a_2E^2 + a_3E^3 + a_4E^4 + \dots$$

$$E_i = \frac{1.5a_4}{a_2} E_\Delta^2 \times E_{2\Delta}$$

when $E_\Delta = E_{2\Delta}$

$$E_i = \frac{1.5a_4}{a_2} E_u^3$$

E_i (DB) = K(DB) + $3E_u$(DB).

Experimental values for 6BH6 tube (E_c = -3, E_s = -130 V):
a_0 = 0.36
a_1 = 0.637
a_2 = 0.64
a_3 = 0.107
a_4 = 1.28.

Computed values of intermodulation for equivalent on channel interference of 1 microvolt: 77 db.

Fig. 10 - Heterodyne converter intermodulation.

$$i_p = a_0 + a_1 E + a_2 E^2 + a_3 E^3 + a_4 E^4 + \cdots$$

$$E_1 = \frac{.75a_4}{a_2} \times E_{L.O.} \times E_u^2$$

WHERE :

E_u IS UNDESIRED SIGNAL

$E_{L.O.}$ IS LOCAL OSC. SIGNAL

COMPUTED 1/2 IF SPURIOUS FOR 1 VOLT LOCAL OSCILLATOR IS 58 db —

Fig. 11 - The 1/2 IF spurious problem.

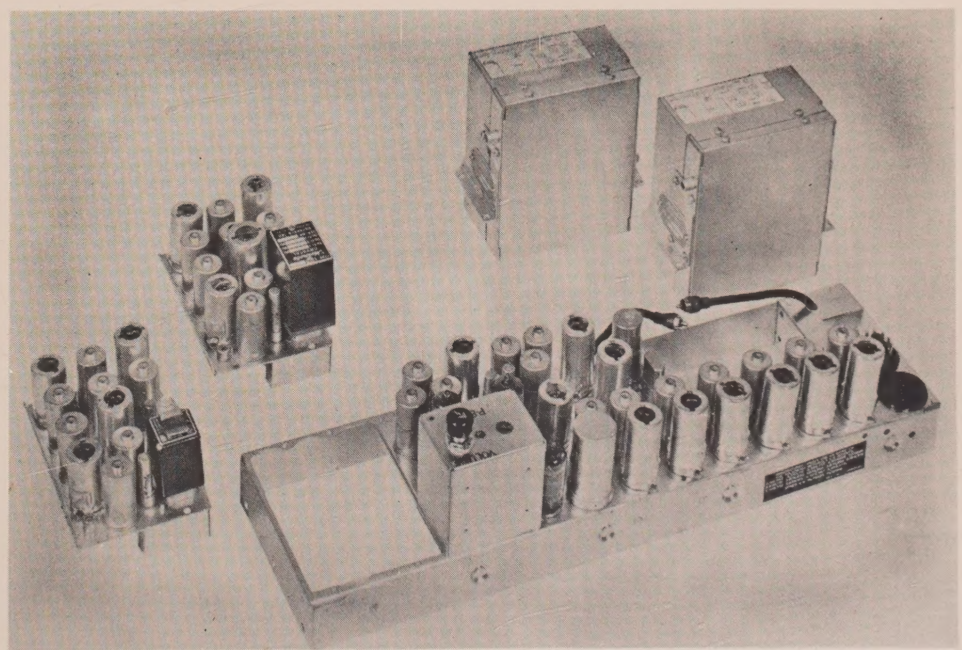


Fig. 12